VI. SOLID-STATE MICROWAVE ELECTRONICS*

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A. SMITH CHART PLOTTING CIRCUIT

Figure VI-1 shows a waveguide circuit that produces an oscilloscope display of the complex reflection coefficient of a load, at a single frequency, as a function of some internal parameter such as diode bias.

The waves reaching the two square-law detectors are superpositions of the incident and reflected waves on the line. The voltages developed in the detectors are therefore the following.

$$V_{horiz.} = \text{const.}\ \{[1 + A\gamma e^{j(\theta-\delta)}] \cdot [\text{comp. conj.}]\}$$

$$= \text{const.}\ [1 + 2A\gamma \cos(\theta-\delta) + (A\gamma)^2]$$

$$V_{vert.} = \text{const.}\ [e^{-j(n+3/4)\pi} + A\gamma e^{j(\theta-\delta+(n+3/4)\pi)}] \cdot [\text{comp. conj.}]$$

$$= \text{const.}\ \left[1 + 2A\gamma \cos\left(\theta - \delta + \frac{3}{2}\pi\right) + (A\gamma)^2\right]$$

$$= \text{const.}\ [1 + 2A\gamma \sin(\theta-\delta) + (A\gamma)^2],$$

where

$$\delta = \frac{4\pi L}{\lambda_g}$$

A = two-way amplitude attenuation coefficient through the load-side attenuator

$\Gamma = \gamma e^{j\theta}$ = amplitude reflection coefficient of the load (time-varying).

If $A\gamma \ll 1$, then the third term in each of the equations above is negligible, so the horizontal and vertical inputs to the oscilloscope are proportional to the real and imaginary parts of the reflection coefficient, rotated by an angle δ, plus a constant. The

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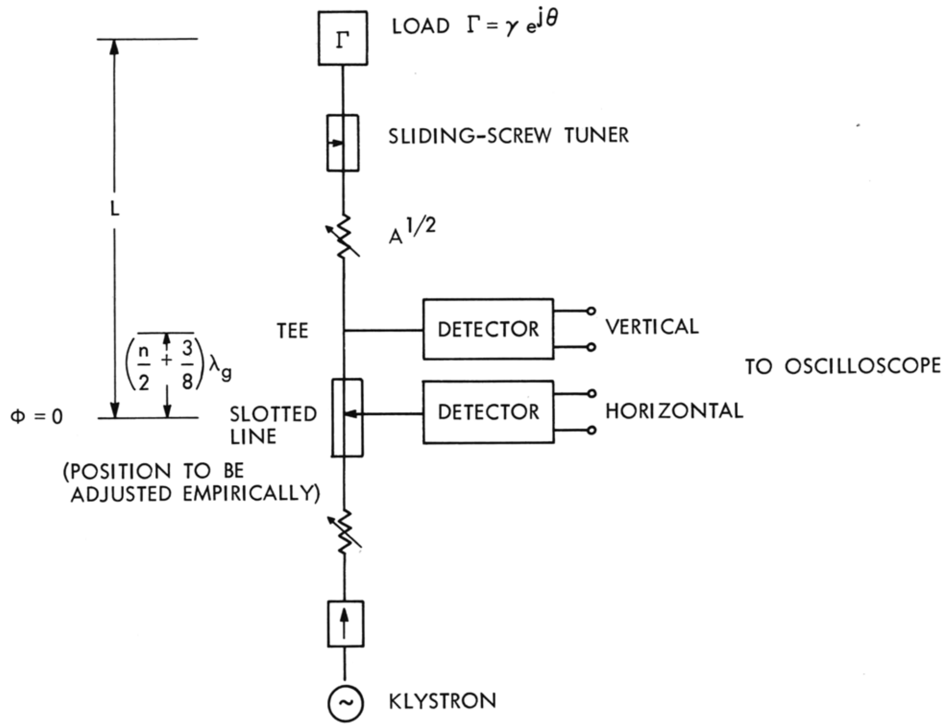

Fig. VI-1. Circuit.

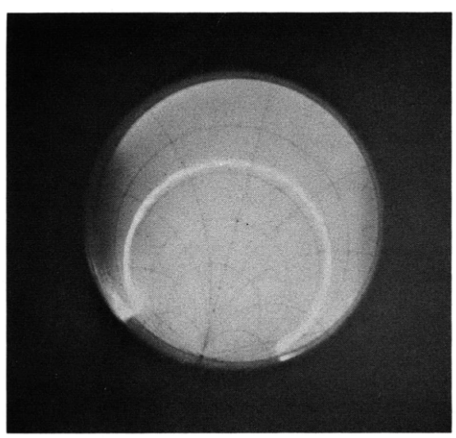

Fig. VI-2. Impedance locus of a parametric amplifier.

constant terms can be taken out with the position controls of the oscilloscope, and the gain controls can be used to counteract any asymmetry in the two halves of the circuit. The origin can be located by setting the load attenuator to maximum, and then with the probe of the sliding-screw tuner run in to maximum VSWR, the load attenuator can be adjusted so that the unit circle of the Γ plane coincides with the edge of the oscilloscope face.

Reflections at the tee and the slotted-line probe have been neglected in this analysis, but since the reflection coefficients of these components are not variables, and multiple reflections from the load are diminished by the attenuator next to it, this is compensated for by adjusting the position of the slotted-line probe so that the unit circle in the Γ plane, as indicated by the sliding screw tuner, maps into a circle on the oscilloscope face.

It is also possible to use a "magic tee" instead of a slotted line in this circuit; in this case, either the E-plane or the H-plane arm would be connected to the detector, and the other arm would be terminated with an adjustable short.

This circuit has proved to be useful in the construction of parametric amplifiers and varactor doublers. In the case of an amplifier, the noise figure is determined by the real axis intercept of the impedance locus as the diode bias is swept from forward conduction to the breakdown point, and the maximum gain is related to the end points of this locus.[1]

Figure VI-2 shows such a locus for a 22.2-GHz parametric amplifier. Here the load attenuator is set at ~10 dB, for 10% accuracy.

P. W. Rosenkranz

References

1. K. Kurokawa, "On the Use of Passive Circuit Measurements for the Adjustment of Variable Capacitance Amplifiers," Bell System Tech. J., Vol. XLI, No. 1, p. 361, January 1962.

B. NONLINEAR CIRCUIT ELEMENTS

A class of nonlinear electrical circuit elements was examined in order to determine which of them are passively realizable and which are not. The class of nonlinear elements examined is defined as follows: Let v_p be the p^{th} time integral of the voltage across the element when p is positive, and the p^{th} time derivative of the voltage when p is negative. Let i_q be the q^{th} time integral of the current through the element when q is positive, and the q^{th} time derivative when q is negative. Then each element in the class under consideration is defined either by the relationship $v_p = f(i_q)$, or the inverse relationship $i_q = g(v_p)$, and the element can be identified by giving the values of the ordered pair (p, q).

Such elements can be divided for my purposes into reactances, for which $p-q$ is odd, and dissipative elements, for which $p-q$ is even. For each of these two groups, frequency-power formulas are derived relating the powers flowing in the element at various frequencies. These formulas reduce to the Manley-Rowe equations for $p + q = 1$, and to Page's and Pantell's inequalities for $p = q = 0$. In their general form they are useful in showing that certain of the elements defined above are not passively realizable, by virtue of the fact that power is not conserved in the element. For nonlinear elements, I have shown that the following are not passively realizable:

a. Elements for which $p - q$ is odd and $p \neq 1$.

b. Elements for which $p - q = 2n$, where n is odd, and $p + q \neq 0$.

For linear elements of this class, the frequency-power formulas do not provide any information about realizability. An analysis of the linear elements can be carried out in the time domain, however, and I have shown that the following linear elements are passively unrealizable:

All those linear elements for which $|p-q| > 1$.

J. G. Webb, Jr.

C. INTERMODULATION DISTORTION

Recent work has been concerned with building equipment capable of extending the dynamic range of spectrum analyzers in the analysis of third-order intermodulation distortion. The measurement of third-order products more than 70 dB below the level of the primary signals is usually meaningless, because of the limited dynamic range of the spectrum analyzer. The attack on this problem is based on cancellation of the primary signals without altering the distortion products.

Prior efforts in the construction of this nulling device have centered on the design and construction of the summing, power-dividing, isolating, and nulling hybrids, P-I-N diode attenuators, power supplies, and coaxial signal-directing switches.

Recent tests have shown that the signal nulling ability is approximately 30 dB wideband and up to 90 dB narrow-band. In the narrow-band case this instrument extends the dynamic range of the 70-dB spectrum analyzer to 160 dB. The intermodulation distortion products were measured on a P-I-N diode attenuator with two -13 dBm signals (at frequencies of 30.000 MHz and 30.001 MHz) applied. The intermod products were found to be 110 below the primary signals.

Improvements on the performance of the test set are planned to increase the wideband cancellation to a minimum of 40 dB and improve the shielding and isolation to values in excess of 150 dB. A slight modification of the system's P-I-N diode attenuators has to be made to give them more range.

R. D. Mohlere

D. AVALANCHE DIODE ANALYSIS

The small-signal, or incremental, impedance of an avalanche diode junction has been measured as a function of the direct avalanche current. These measurements were made in the frequency range 4-12 GHz in increments of 1/2 GHz.

At a given value of bias current, the frequency variation of the incremental impedance is very similar to that of the circuit shown in Fig. VI-3.

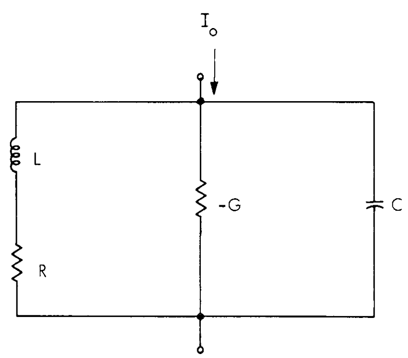

Fig. VI-3. Avalanche diode junction equivalent circuit.

The frequency-independent circuit elements (R, L, G, C) are chosen so as to yield a minimum value for the error expression

$$E = \frac{1}{N} \sum_{i=1}^{N} \left| Y_{mi} - Y_{ci} \right|^2,$$

where Y_{mi} is the measured value of admittance at frequency f_i, Y_{ci} is the admittance of the circuit at frequency f_i, and the sum extends over the measuring frequencies.

Following this procedure for several values of bias gives the dependence of the elements on the direct current I_o.

$$L \propto \frac{1}{I_o}$$

$$G \propto I_o$$

$$R \propto \text{const.}$$

$$C = C_{min} = \text{depletion layer capacitance at avalanche breakdown.}$$

Further analysis will be directed toward a large-signal model for the diode.

D. F. Peterson